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Electrically Small Complementary Pair (ESCP) With Inter-Element Coupling

Rentry Systems Division
Development Operations
The Aerospace Corporation
El Segundo, Calif. 90245

10 December 1976

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Electrically small (reduced size) antennas are inherently narrowband, or inefficient, or both. A summary is presented of prior work on complementary pair antennas and the use of the coupling between the two elements to optimize the impedance match and efficiency of such structures. The design of electrically small complementary pairs is described, and preliminary measurement results are shown for monopoles. These measurements indicate a substantial improvement in gain-bandwidth product as compared to conventional matching techniques for electrically small antennas.

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In addition, the complementarization circuit in conjunction with the hybrid feed can reduce the effects of mutual impedances in steerable-beam arrays. Applications include small individual antennas and scatterers as well as linear, planar, or circular arrays in a variety of fixed-station or vehicular uses.

Various types of decoys and ECM systems for aircraft and missiles are examples of potential applications of electrically small antennas and scatterers.

Large, hardened, ground-based phased arrays and wideband communication base station antennas are typical phased-array applications in fixed stations. Vehicular applications encompass wideband direction finders, aircraft conformal antennas for various functions, and high-performance spacecraft antennas for communications and radar systems.

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FIGURES

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I. INTRODUCTION

Whenever available installation height is limited, the antenna can be foreshortened so as to fit into the limited space. This causes the antenna impedance to become very reactive. Past practice was to tune out the capacitive reactance by means of an inductor. This renders such an "electrically small" antenna narrowband, and its efficiency is reduced by losses occurring in the tuning circuits. This problem becomes substantial for dipole lengths of less than $\lambda/8$, which is generally used as the criterion for electrically small antennas. Typical performance results are: 1 to 10 percent bandwidth and 70 to 30 percent efficiency, respectively, for a monopole height of 0.05 wavelengths [1].

An alternate approach for tuning a short dipole or monopole consists of using two of the antennas, which are mutually coupled, and matching the input reactance of one with the reactance of the other after it has gone through an inversion circuit. This inversion circuit is realizable in the form of an externally complementarized hybrid feed circuit similar to the one described previously for resonant-height antennas [2]. Mutual coupling between the two elements in the pair can be adjusted in a constrained design volume by varying (a) the length-to-diameter ratio of the elements, and (b) the element spacing and feed cable length differential for phasing.

II. THE COMPLEMENTARY PAIR MATCHING PRINCIPLE

The "Complementary Pair Element Group" or CPEG was first conceived as a broadband element group with basically resonant radiator length. The radiators were placed in a phased array environment, or in front of a reflector, or both. In both cases detuning of the input impedance resulted for a single dipole (or monopole) due to mutual coupling, either with adjacent elements in the array, or with the image element(s) created by the reflector, or both. The solution sought was a matching system that would (a) match out individual element mismatches caused by operation over a wide frequency range (without having to resort to electrically large structures such as log-periodics), and (b) maintain this match independent of the above cited mutual coupling effects.

Two basically different approaches, briefly described in [2] and [3], were investigated: self-complementary pairs and pairs of identical radiators where complementarity was achieved by modifying the impedance of one radiator through a "complementarizing" network, such as a delay line. In both cases, the impedance averaging properties of 180-deg hybrid tees were used [4] to match out the resulting sets of complementary impedances seen at the hybrid output ports (Fig. 1). (Note that the symbol $\underline{|*}$ is used to describe complementarity.) Figure 2 shows examples of different complex complementary loads, each pair leading to a perfectly (within the hybrid design) matched sum port, at the expense of power loss in the difference port load equal to that suffered in a load isolator.

For the purpose of achieving optimum impedance match under all beam steering conditions and over wide bandwidth in a phased array environment, it soon became apparent that self-complementary elements placed side by side would not be practical either from a grating lobe standpoint or because of mutual coupling [5]. Figure 3 shows the basic feed circuit for an externally complementarized endfire pair, and Fig. 4 is typical embodiment

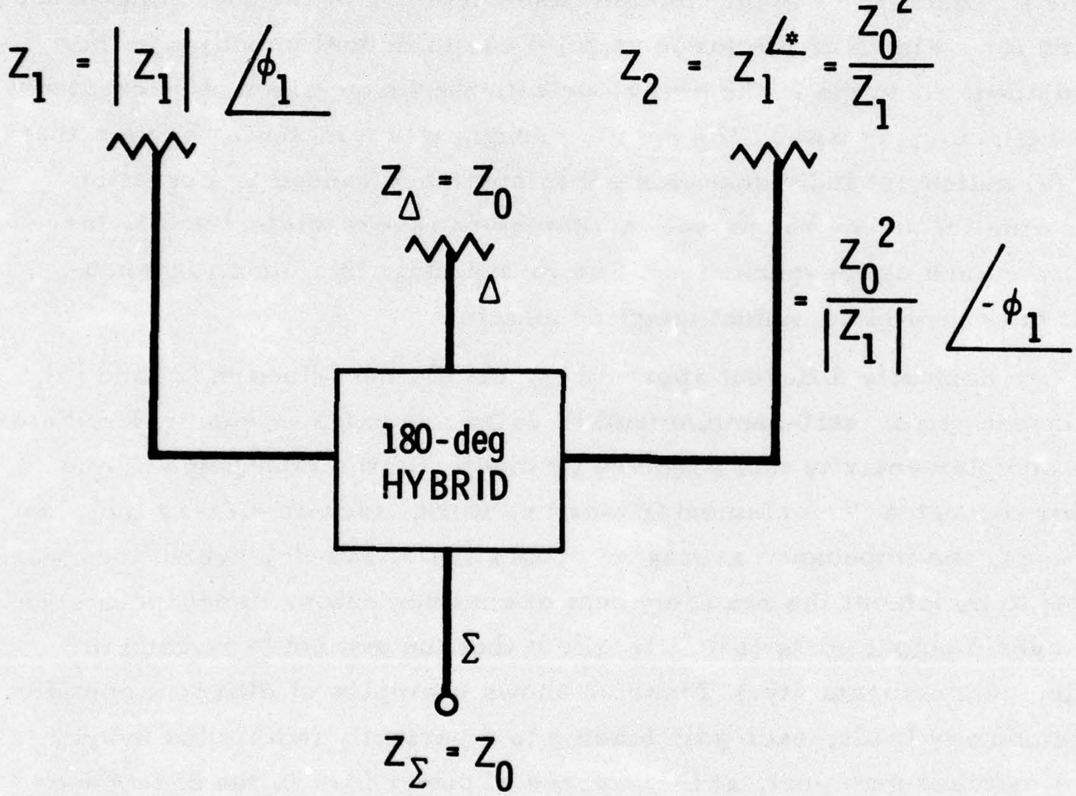

Fig. 1. General Impedance Relations

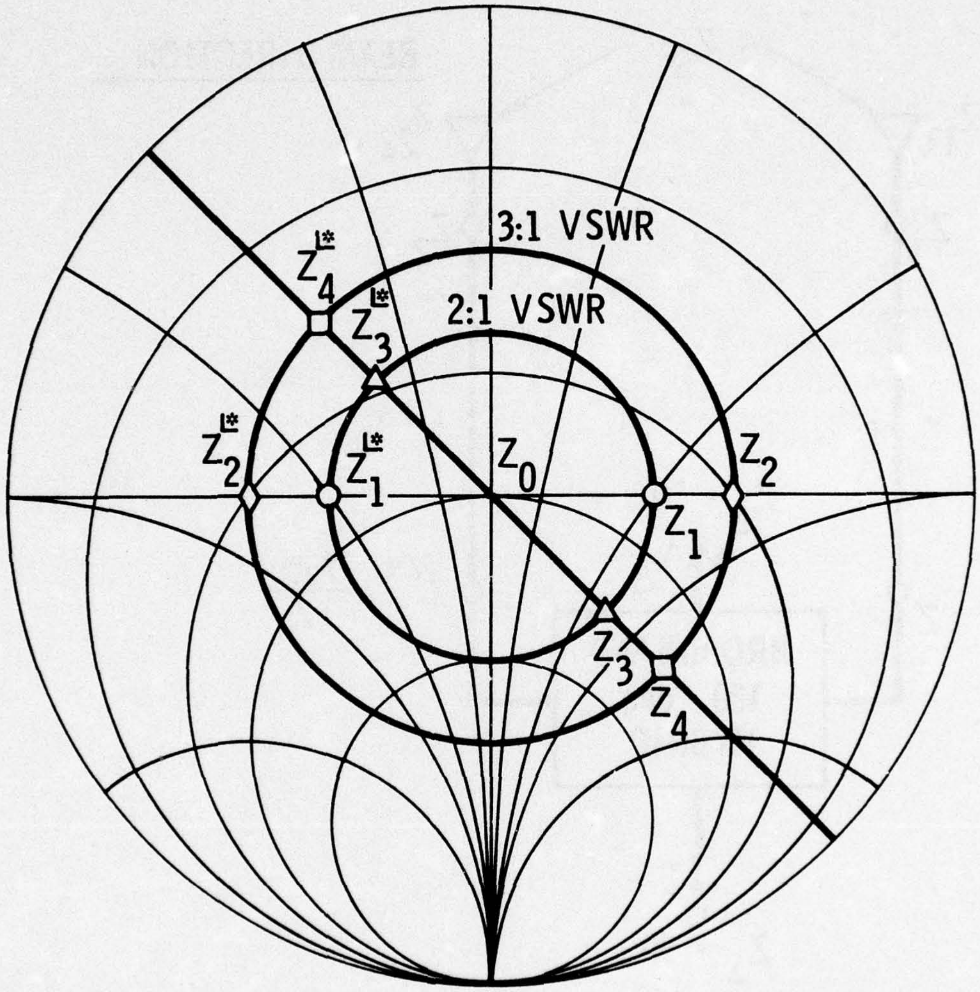

Fig. 2. Arbitrary Complementary Loads

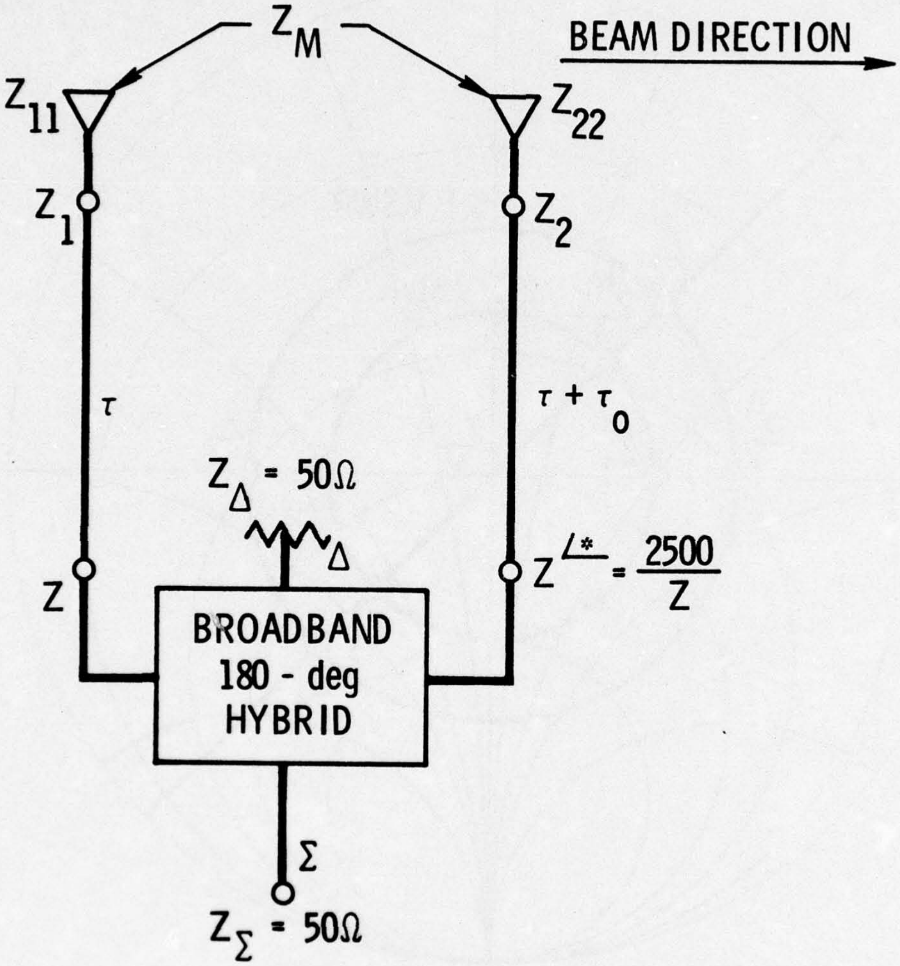

Fig. 3. Broadband Endfire Complementary Pair Feed Circuit

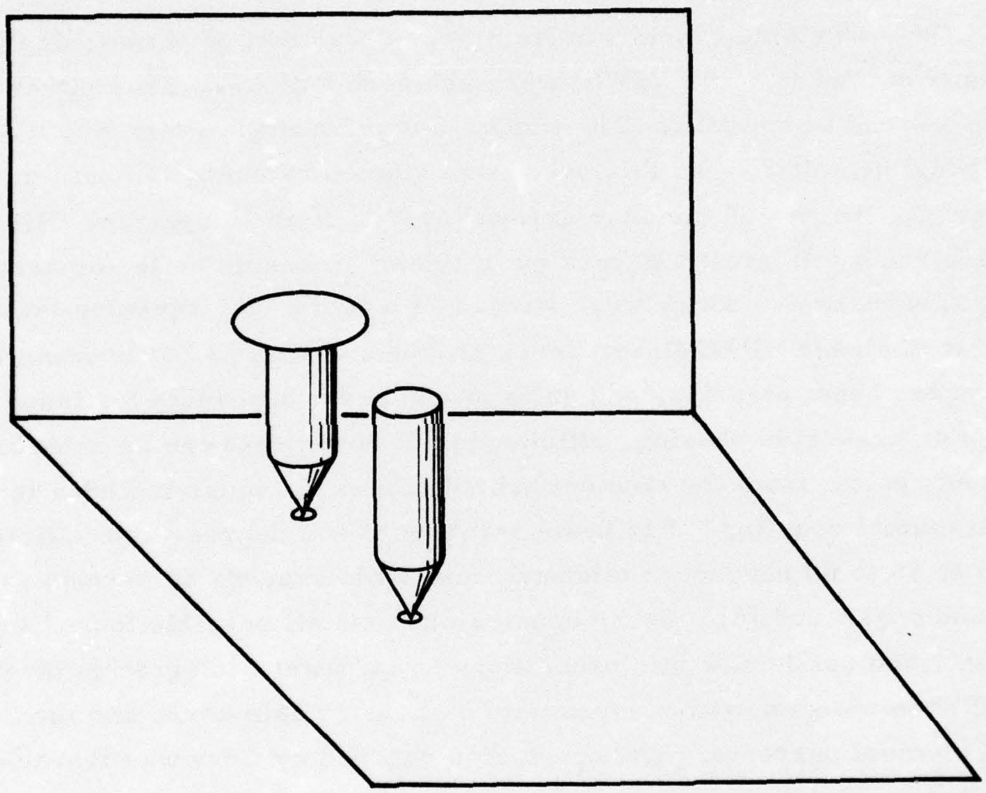

Fig. 4. Endfire Complementary Pair Element Group With Reflector

of a monopole pair in front of a reflector. (The reflector was added to the array to provide maximum front-to-back ratio for purposes of interference suppression.) Figures 5 and 6 show measured monopole impedances and Fig. 7 shows the VSWR for the group input at the hybrid sum port. Figure 8, which is the power measured at the different port, represents matching loss other than hybrid insertion losses [2], [3], [5].

The surprising conclusion from Fig. 8 was that good matching properties (from 6:1 to 1.5:1 VSWR) were achieved with very little power loss at the low end of the band. The endfire complementary pair, which emerged as a possible solution for this case, was studied extensively, both in the form of linear and circular arrays [6] [7]. Both an enlarged UHF version and a full-scale HF version of linear broadside reflector arrays using endfire pairs, which were studied as a follow-on, revealed that: "...A suitably designed CPEG linear array provides at least 3.5:1 bandwidth for ±45 degree beam steering, and quite probably 4:1 bandwidth for lesser steering. For broadside phasing, efficiencies of 80 percent can be achieved for the feed system from the element hybrids forward, which includes losses due to mutual coupling. For beam steering to ±45 degrees, the efficiency drops to 55 to 60 percent at midband, and again exceeds 80 percent towards the band edges..." [8]. These conclusions took all possible losses into account, and confirmed gain predictions from pattern integration through actual absolute gain comparisons with a standard gain horn, and monitoring of all element currents. The conclusion reached by Cory was that the impedance matching was "phenomenally good" and was due to some "fortuitous averaging in the hybrids". The conclusion reached by this author is the realization that inter-element coupling within the pair can be made to enhance the matching process at the low-frequency end of a typical 3:1 frequency range, and that mutual coupling effects due to adjacent pairs in a phased array can be minimized. Both properties are inherent to the hybrid feed circuit and to the external complementarization process. The maintenance of an optimum amount of endfire coupling thus emerged as one of the

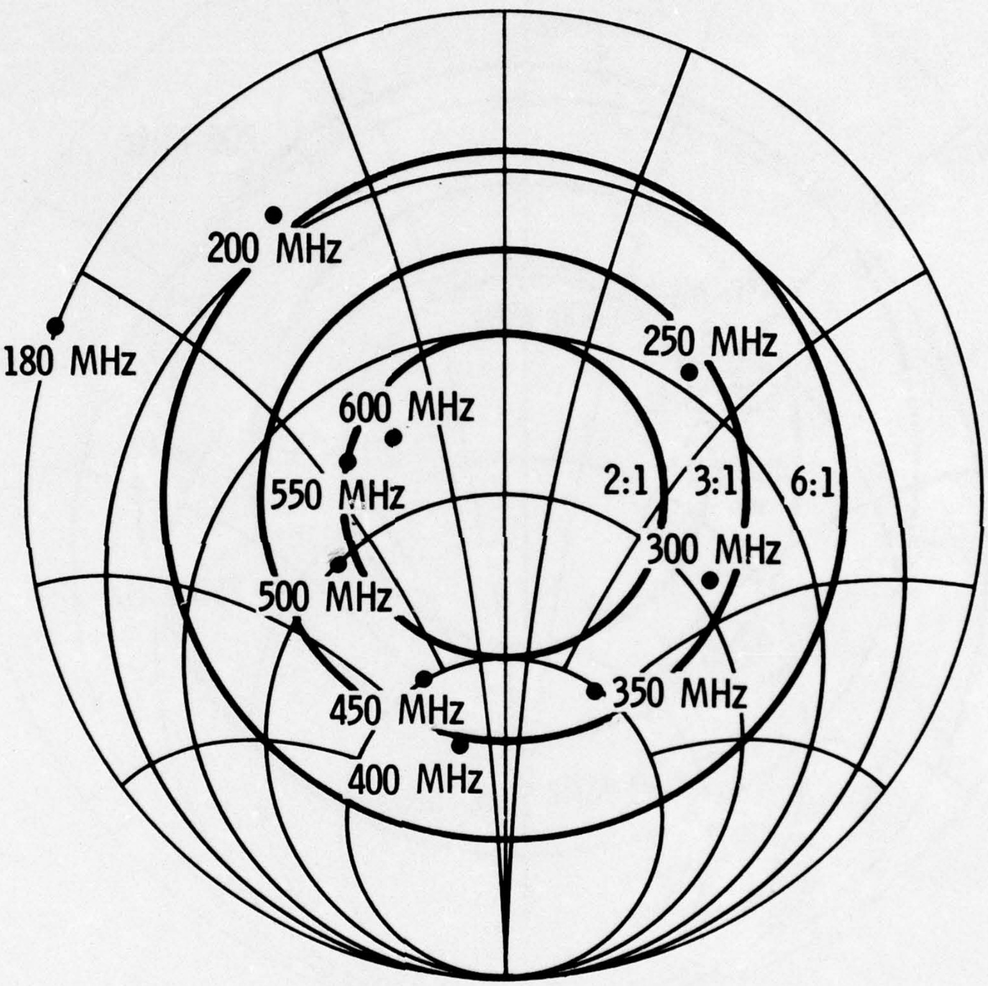

Fig. 5. Endfire Monopole CPEG Front Element Impedance ($Z_o = 50\Omega$)

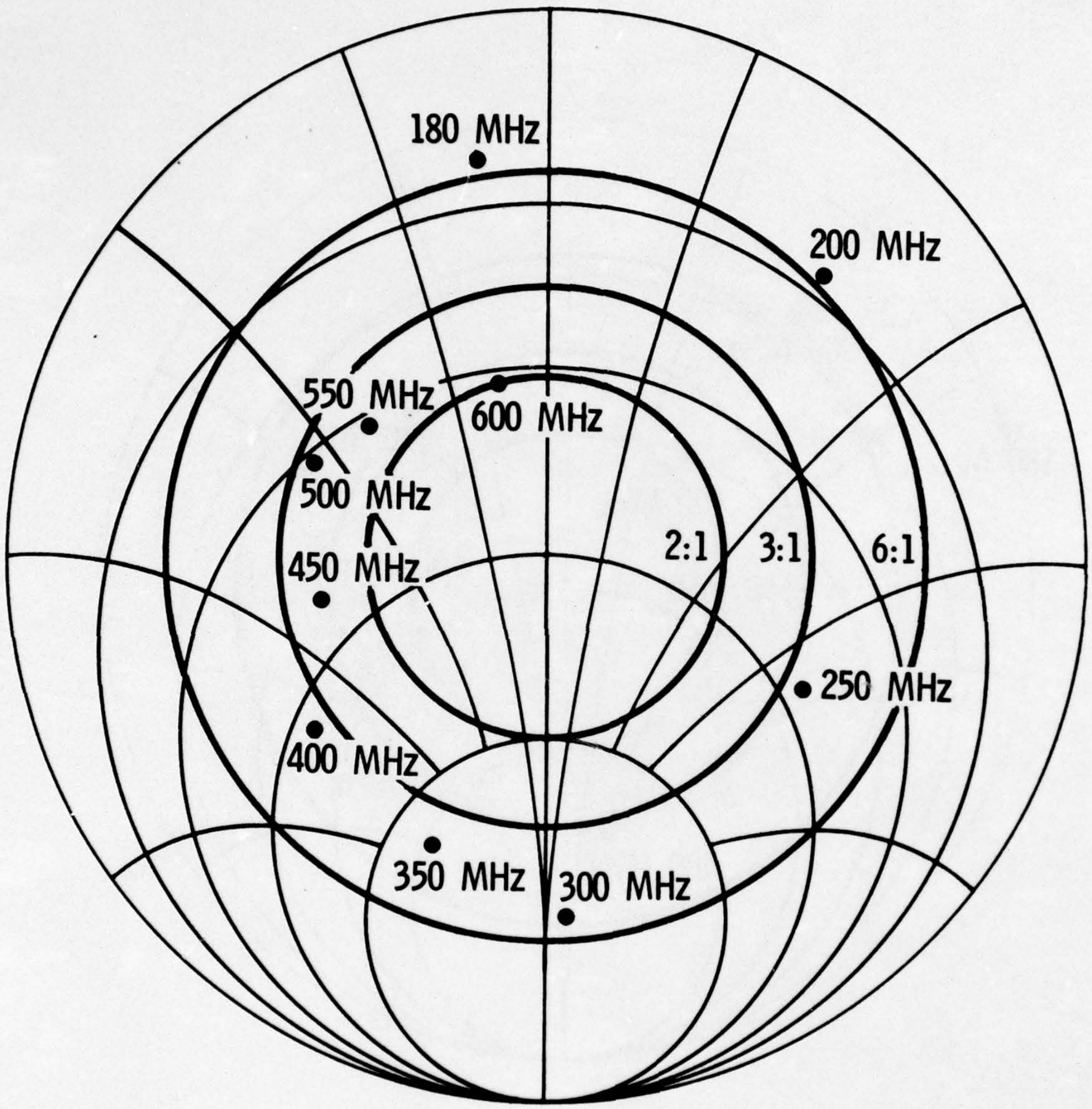

Fig. 6. Endfire Monopole CPEG Back Element Impedance ($Z_o = 50\Omega$)

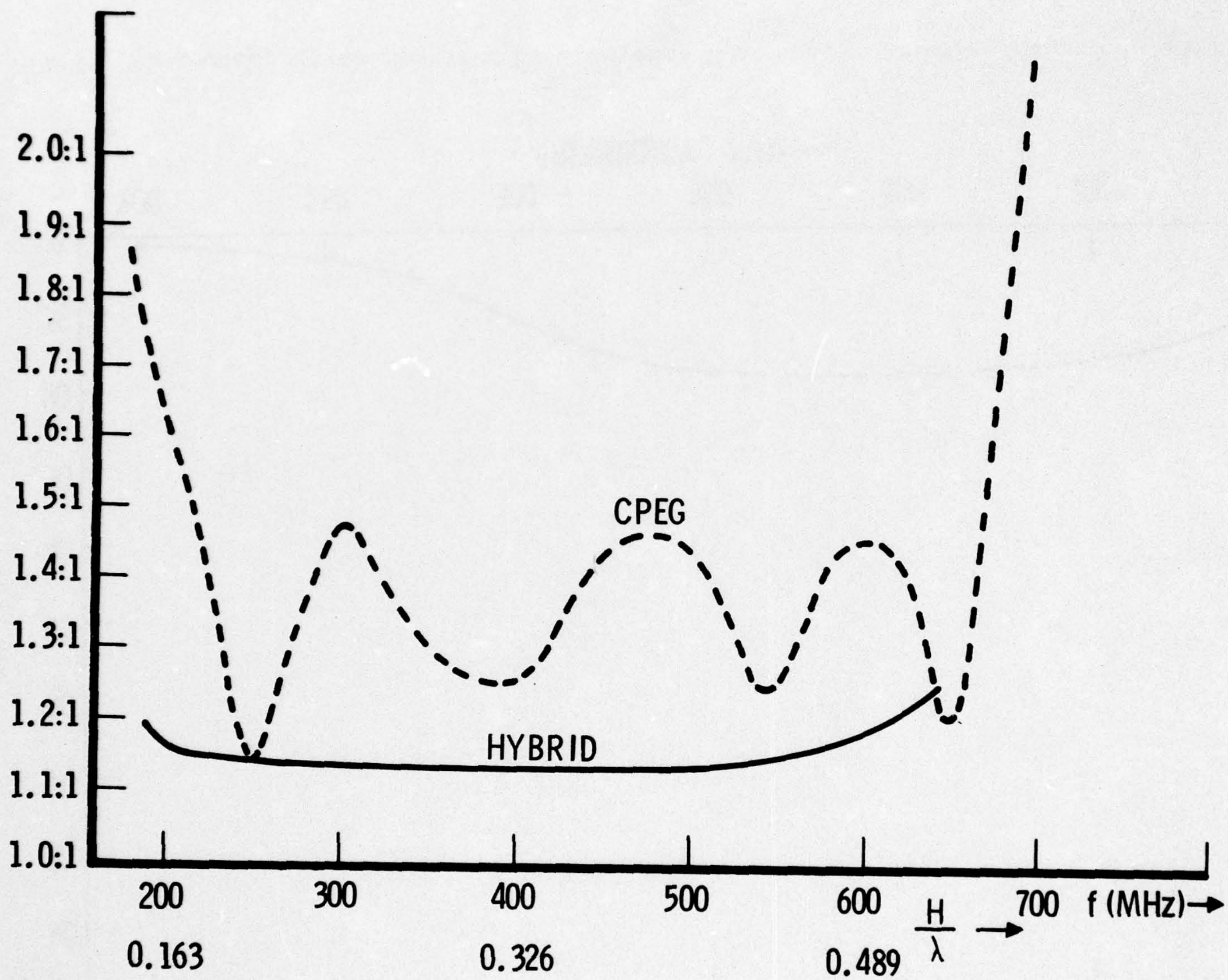

Fig. 7. Standing Wave Ratio of Endfire Monopole CPEG

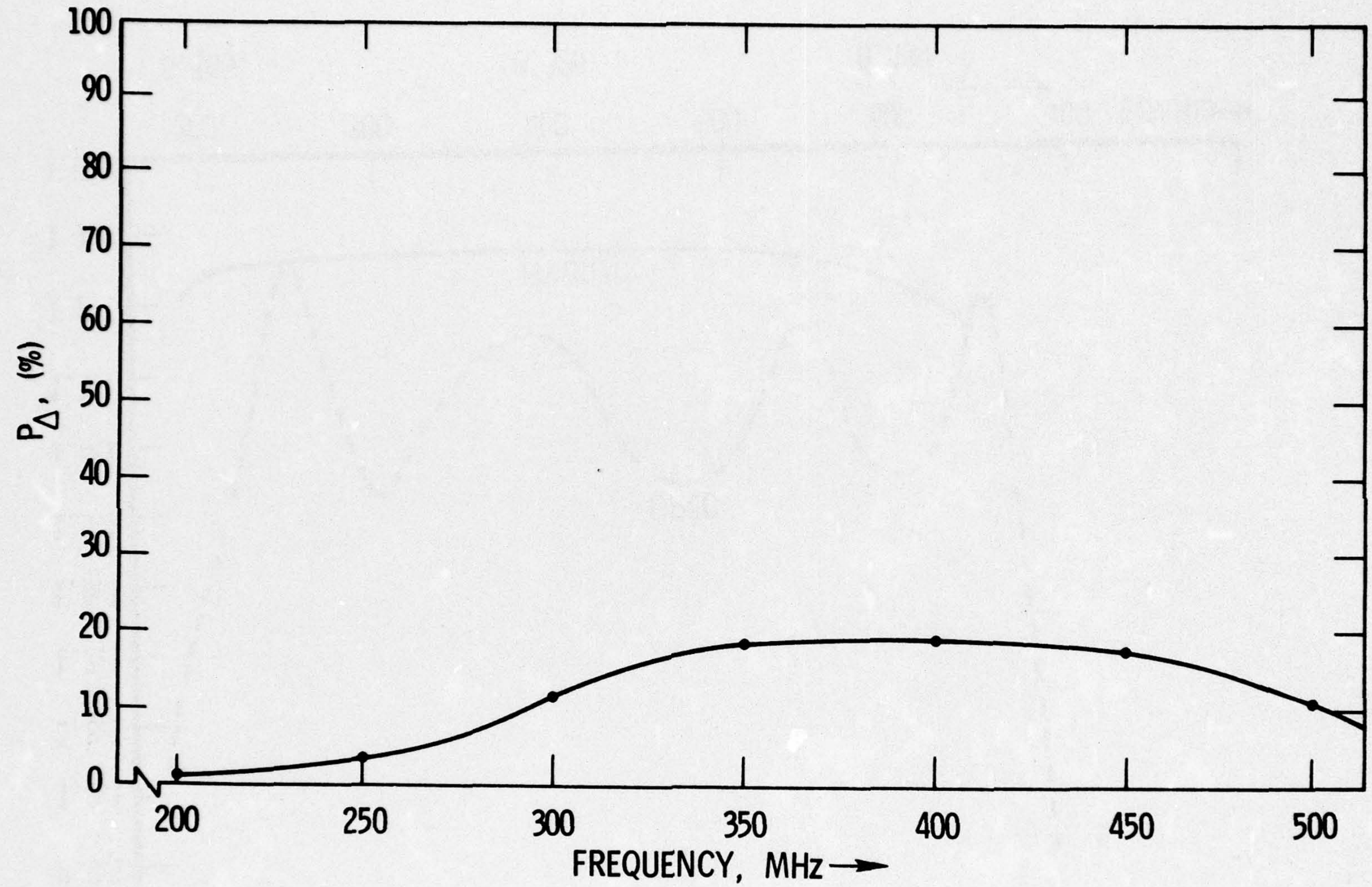

Fig. 8. Percentage Power Absorbed by Difference Port Load for Endfire Monopole CPEG

design criteria for endfire complementary pairs, when the radiators were approaching the electrically small regime. On the other hand, the straightforward combination of self-complementary elements (a slot and a monopole, specifically) had been shown to suffer from very low efficiency when the radiators were electrically small, partly because of power lost in the load terminating a feed line [9].

III. THE ELECTRICALLY SMALL COMPLEMENTARY PAIR (ESCP)

Since inter-element mutual coupling is strongly dependent on the electrical height of the radiators, and the only other variables are the fatness and shape of the elements, it is intuitively obvious that a reduction in coupling due to reduced height has to be compensated for by vastly decreasing the length-to-diameter ratio. To determine the minimum useful "fatness", several design shapes were studied experimentally, starting with some relatively thin elements. (The length-to-diameter ratio of the resonant-height monopoles described in [2] through [8] was about 3, with a 60-deg cone angle at the base.) A design configuration considered acceptable regarding VSWR at the lowest frequency is shown in Fig. 9. The total height of the radiators was one eighteenth of a wavelength at the lowest frequency, the height of the 90-deg conical section being exactly half the total height for a length-to-diameter ratio of unity. The measured VSWR of this ESCP configuration is shown in Fig. 10, in comparison to the VSWR of an isolated monopole of the same dimensions. ESCP complementarity was maximized at λ_o through a quarter-wave length of transmission line.

To evaluate the total matching loss of this ESCP configuration, the power absorbed in the difference port load of the hybrid was measured as shown in Fig. 11. The red paths illustrate the power flow. Of the total power P_o entering the sum port, a certain percentage, reduced by sum port reflection and hybrid insertion loss, reaches the elements. There, a portion gets reflected; the rest is radiated except for a small portion accepted by the other element and re-channeled into the circuit. The major advantage of the mutual coupling apparently lies in an adjustment of the element reactance. Indeed, measurements during the programs reported in [5] and [8] have shown that not only the sum port VSWR improves, but the individual element reflections drop drastically as soon as the loop is closed by

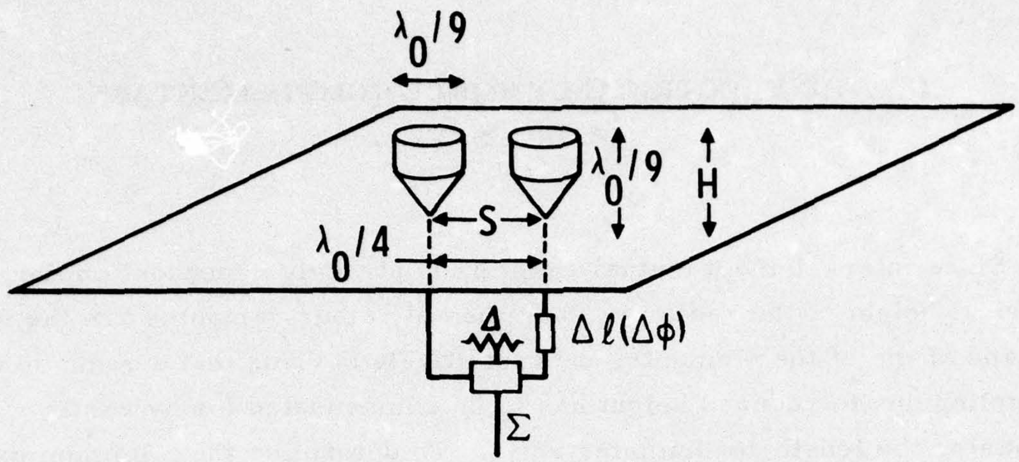

Fig. 9. ESCP Test Configuration

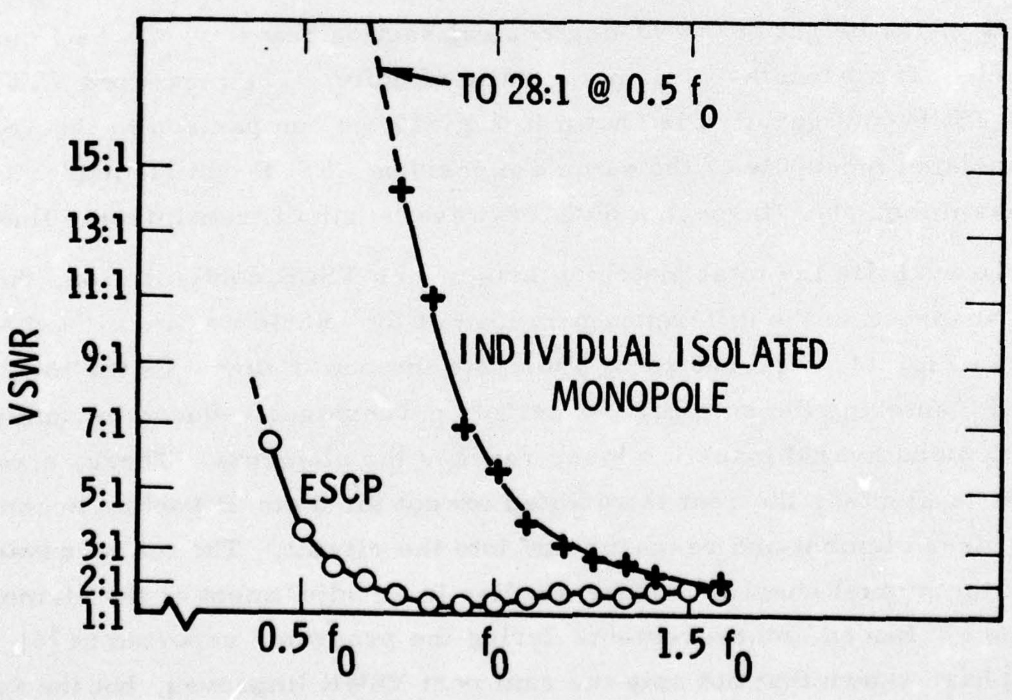

Fig. 10. ESCP Input Impedance Match

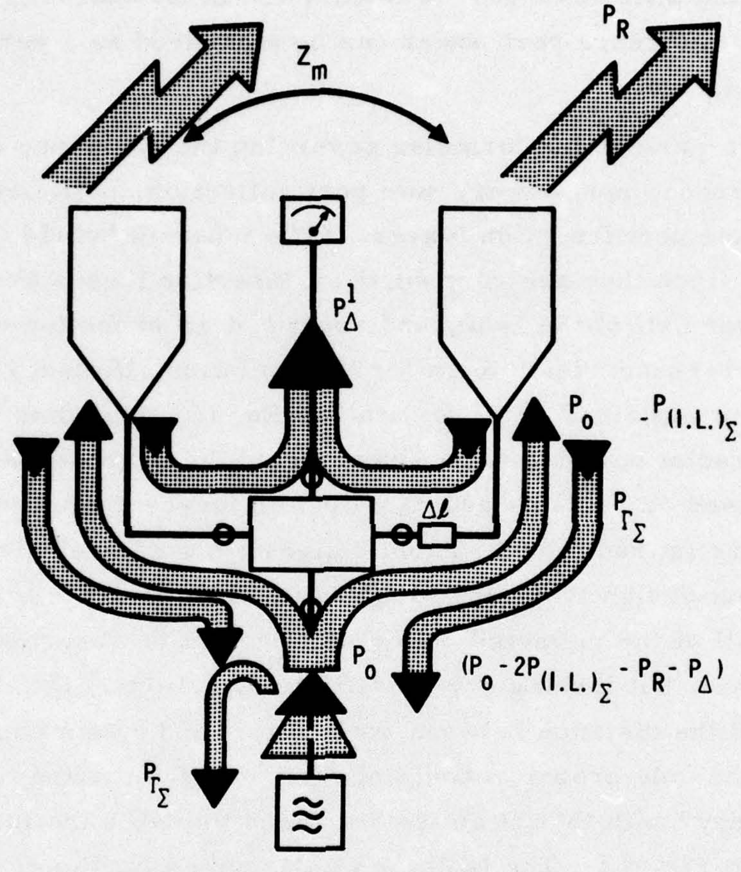

Fig. 11. Power Flow Diagram for ESCP Matching Loss Measurements

the hybrid. The power reflected from the elements gets divided into the sum port and the difference port. The sum port portion has already been accounted for and can simply be determined from the sum port reflection coefficient when the difference port is terminated in its characteristic impedance. The difference port power can be measured as a percentage of P_o as shown in Fig. 11.

Table 1 summarizes the formulas governing the efficiency calculation based on the difference port power, sum port reflection coefficient, and sum-and-difference port insertion losses. (The inherent hybrid mismatches can be neglected since they are very small.) Insertion losses are less than 0.5 dB in the lower half of the band, and about 0.4 dB at the lowest frequency. Equation (1) includes a term for the radiation efficiency of the elements. This is explained in more detail in Eq. (5), including an inequality for the range of radiation resistance depending on the current distribution. The efficiency based on radiator and cable ohmic losses can usually be neglected for these fat radiators. Table 2 gives a sample calculation for the ESCP at the lowest design frequency, leading to -4.8 dB. It is assumed in this calculation that all of the reflected power will have to be absorbed. This may be pessimistic, but lacking any specific information on the transmitter output circuit and the distance between transmitter and antenna in wavelengths, this is the only proper accounting method. Comparing this "Total Matching Efficiency" with that of an isolator used to match the individual monopole leads to Fig. 12. The isolator was assumed lossless, lacking any specifics. Any particular isolator (if available at, say, VHF) may have a substantial insertion loss, of course, which will have to be added to the values of Fig. 12.

Actual absolute gain, in both cases, would have to be determined from gain measurements on a pattern range. However, the previous laborious procedures described in [8], which failed to identify any "hidden loss"

Table 1. ESCP Efficiency Formulas

$$\eta_{tot} = \eta_{\mu} \times \eta_{\Sigma} \times \eta_{R} \times 100 \qquad [\%] \tag{1}$$

with

$$\eta_{\mu} = 1 - \log^{-1}\left\{\log \frac{P_{\Delta}}{P_o} - \left[\frac{(I.L.)_{\Sigma}}{10} + \frac{(I.L.)_{\Delta}}{10}\right]\right\} \tag{2}$$

$$\eta_{\Sigma} = \eta_{(I.L.)_{\Sigma}} \times \eta_{\Gamma_{\Sigma}}$$

$$= \log^{-1}\left[\frac{(I.L.)_{\Sigma}}{10} + \log\left(1 - \Gamma_{\Sigma}^{2}\right)\right] \tag{3}$$

Γ_{Σ} = True Sum Port Reflection Coefficient referred to element VSWR

$$= - \frac{(VSWR)_{el} - 1}{(VSWR)_{el} + 1} \tag{4}$$

with $\mathrm{Coth}^{-1}(VSWR)_{el} = \mathrm{Coth}^{-1}(VSWR)_{m} - \frac{\alpha}{8.686}$

and

$(VSWR)_{el}$ = element standing wave ratio

$(VSWR)_{m}$ = measured (hybrid) VSWR

α = hybrid attenuation constant

Table 1. ESCP Efficiency Formules (cont.)

$$\eta_R = \frac{R_R}{R_R + R_A + R_C} \quad (5)$$

Where

R_R = Element Radiation Resistance

$$40\,\pi^2 \left(\frac{h}{\lambda}\right)^2 < R_R < 160\,\pi^2 \left(\frac{h}{\lambda}\right)^2$$

and

R_A = Element Ohmic Losses

R_C = Cable Ohmic Losses

Typically,

$$\eta_R \geq 90\% \qquad \text{for } h \geq \frac{\lambda}{18}$$

Table 2. ESCP Efficiency Budget

$$f = \frac{1}{2} f_o$$

Hybrid Insertion Losses	$(I.L.)_\Sigma$	=	0.4 dB	Sum Port
	$(I.L.)_\Delta$	=	0.4 dB	Difference Port
Corrected Sum Port Reflection Loss (3.8:1 VSWR) $\Gamma_\Sigma = 0.583$		-	1.8 dB	
Power Measured at Difference Port		-	4.3 dB wrt Σ input	
Actual Difference Port Power		-	3.5 dB wrt Σ input	
Difference Port Matching Efficiency	η_μ	=	$1 - \log^{-1}$ (-3.5 dB) = 55%	
Sum Port Reflection Efficiency	η_{Γ_Σ}	=	66%	
Sum Port Insertion Loss Efficiency	$\eta_{I.L._\Sigma}$	=	91%	
Total Efficiency (Excluding Radiator, Cable Losses)	$\frac{\eta_{tot}}{\eta_R}$	=	$0.55 \times 0.66 \times 0.91$	
		=	0.33	
		=	-4.8 dB	

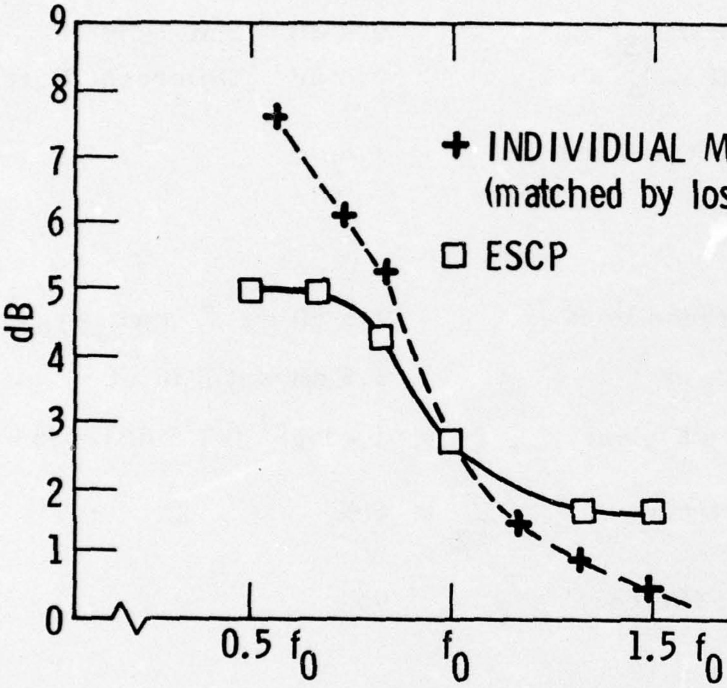

Fig. 12. ESCP Matching Loss (Assuming all of reflected power of Fig. 2 is converted into loss)

mechanism, should justify the assumption that for the ESCP, the net absolute gain can be roughly estimated from the individual small radiator pattern gain (1.75 dB + 3 dB for ground plane) plus the pair directivity gain as illustrated in Fig. 13 [10], minus the matching losses given in Fig. 12. The total gain bandwidth product thus appears to have been improved with respect to the classical inductor-matched electrically small monopole antenna. The matching is completely instantaneous, so that wideband signals can be transmitted. Using a definition of bandwidth where the minimum radiator dimension and the minimum efficiency occur at the lowest operating frequency (as shown in Figs. 10 and 12), we can assign a bandwidth of 3:1, or relative bandwidth of 100 percent where an octave equals 66 percent. The minimum efficiency is approximately 25 percent, or -6 dB, and the directive gain in the direction of propagation approximately 5.75 dB. Hence, unity gain/bandwidth product exists, as compared to 0.1 G × B for a 30 percent efficient monopole with "broadband" matching to 10 percent relative bandwidth, or compared to 0.02 G × B for a monopole tuned to maximum efficiency and minimum bandwidth. It appears, therefore, that about an order of magnitude improvement in gain X bandwidth product should be feasible.

Potential applications of the ESCP include directional elements for scanning wideband phased arrays, small antennas for low-silhouette requirements on various vehicles, and other applications calling for wideband antennas or scatterers with directional pattern over part of the band.

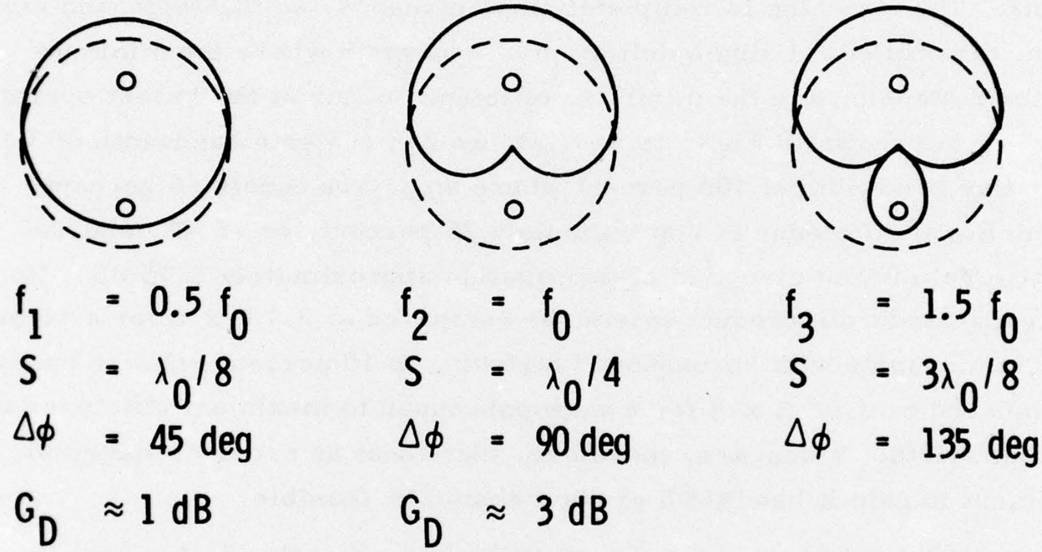

Fig. 13. ESCP Pattern Summary

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